Dual-Polarization Transmit-Receive Reflectarray Antenna Made of Cells with Two Orthogonal Sets of Coplanar Parallel Dipoles

R. Florencio[(1)], J. A. Encinar[(2)], R. R. Boix[(1)].
rflorencio@us.es, jose.encinar@upm.es, boix@us.es.
[(1)]Department of Electronics and Electromagnetism. University of Seville. Avda. Reina Mercedes, s/n. 41012-Sevilla.
[(2)]Department of Electromagnetism and Circuit Theory. Polytechnic University of Madrid.
C. Ciudad Universitaria s/n 28040 Madrid.

***Abstract*— A dual–polarization dual–frequency focused beam reflectarray antenna is designed for transmit–receive (Tx–Rx) operation in Ku Band. A novel broadband reflectarray element is introduced, which consists of two orthogonal sets of three parallel dipoles. To adjust the dimensions of the elements in the antenna, we have used the local periodicity assumption, and we have analyzed a multilayered periodic structure surrounding each element by means of a home–made software based on the method of moments in the spectral domain. This strategy makes it possible to design reflectarray antennas within reasonable CPU times. The designed antenna shows a bandwidth of 10.9% in Tx–band and 7.1% in Rx–band for gain variations lower than 1 dB. Also, levels of cross–polar components 30 dB below the radiation maximum have been achieved. In order to improve the cross-polar discrimination, the elements of the designed reflectarray have been slightly rotated, and this has made it possible to achieve levels of cross–polar components 35 dB below the maximum in both the Tx and the Rx band.**

I. Introduction

Printed reflectarrays are planar reflector antennas made of one or more layers of microstrip patch arrays. The microstrip patches of the arrays are conveniently tuned to produce a progressive phase shift of the reflected field. This makes it possible to collimate or to conform the radiated beam when the reflectarray is illuminated by a feed (usually a horn antenna), just as it happens with conventional reflectors [1]. Printed reflectarrays present several technological advantages when compared with conventional reflectors and phased arrays. On the one hand, they are lighter and easier to manufacture than reflector antennas, and they show improved polarization performance [2]. On the other hand, they are simpler and more efficient than conventional microstrip arrays because they eliminate the complexity and the losses of the feeding network [2].

Some of the first prototypes of reflectarray antennas with elements of variable size were fabricated with one single layer of rectangular patches [2]. Unfortunately, this type of element presented some drawbacks such as reduced phase range and reduced bandwidth [3]. All the drawbacks shown by the single layer elements have been overcome by the use of multi-resonant elements [4]. A convenient multi-resonant reflectarray element consists of three coplanar parallel dipoles [5], [6]. This element has demonstrated a bandwidth performance which is similar to that of the element based on stacked patches [7], but the element based on coplanar paralell dipoles has the additional advantage that its manufacturing process is easier and cheaper. However, this element is only useful for single polarization reflectarray. In this work a transmit–receive focused beam reflectarray for dual–polarization dual–frequency applications has been designed by using elements with two orthogonal sets of three parallel dipoles. It has been proven that a slight rotation of the dipoles in the initial design makes it possible to improve the cross–polarization discrimination of the antenna [8].

II. Characterization of Reflectarray Element for Dual–Polarization Dual–Frequency Applications

Fig. 1 shows the reflectarray element considered in this work. This reflectarray element consists of two orthogonal sets of three coplanar parallel dipoles. The dipoles are printed on a substrate comprising two layers of thickness h_i $(i = 1, 2)$ and complex permittivity $\varepsilon_i = \varepsilon_0 \varepsilon_{ri}(1 - j \tan \delta_i)$ $(i = 1, 2)$. The bottom dielectric layer is limited below by a ground plane. The center of the dipoles of length l_{2i} $(i = 1, 2)$ coincides with the center of the unit cell. The dipoles of length l_{1i} are symmetrically placed with respect to the dipole of length l_{2i} $(i = 1, 2)$.

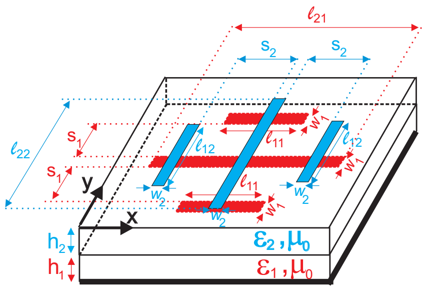
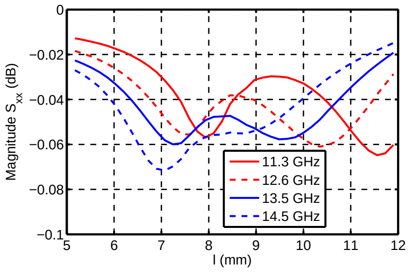
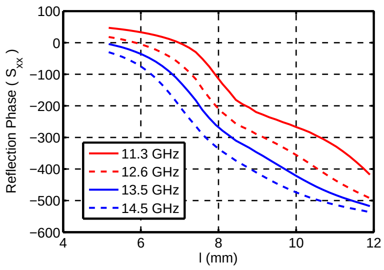


Fig. 1. Reflectarray element cell based on two orthogonal sets of three parallel dipoles.

This unit cell will be the reflectarray element used for the design of a dual–frequency dual–polarization reflectarray antenna. When adjusting the sizes of the dipoles of the reflectarray that lead to the appropriate reflection phases for a focused beam pattern, we will assume that each element is embedded in an infinite periodic array of elements of the same size [2]. This is known as local periodicity assumption [1], and it makes it possible to design reflectarray antennas within reasonable CPU times [2], even for very stringent requirements [9]. The validity of the local periodicity assumption is based on the fact that it leads to numerical results that show good agreement with experimental results [2], [9]. A home-made software based on the method of moments in the spectral domain (MoM-SD) [10] is used for the analysis of the periodic structures required in the design of the reflectarray antenna under the local periodicity assumption. This software has already been validated in [6] by comparison with CST. The home–made software has proven to be between two and three orders of magnitude faster than CST, which is crucial for reflectarray design within reasonable CPU times.



(a)



(b)

Fig. 2. Magnitude (a) and phase in degrees (b) of the reflection coefficient of the reflectarray element of Fig. 1 when it is placed in a periodic environment. Normal incidence and X–polarized electric field are assumed. The results are presented at the limits of the Tx and Rx frequency intervals. Parameters: $a = b = 12$ mm; $l_{21} = l_{22} = l; l_{11} = l_{12} = 0.63l$; $s_1 = s_2 = 3.5$ mm; $w_1 = w_2 = 0.5$ mm; $\varepsilon_{r1} = 2.17$; $\tan\delta_1 = 0.0009$; $h_1 = 3.175$ mm; $\varepsilon_{r2} = 2.17$; $\tan\delta_2 = 0.0009$; $h_2 = 0.127$ mm.

Fig. 2 shows the magnitude and phase of the reflection coefficient of the reflectarray element of Fig. 1 in a periodic environment as a function of the length of the central dipoles. The results are presented for X–polarization at the limits of both the Tx frequency interval (center frequency: 11.95 GHz) and the Rx frequency interval (center frequency: 14 GHz) under normal incidence. For all the frequencies considered, the phase curves show very linear behavior with smooth slopes, the phase range covered is typically 500 degrees, and the losses are lower than 0.08dB. Similar results have been obtained of the reflection coefficient for Y–polarization. This indicates that the element of Fig. 1 is a very adequate reflectarray element [3] for dual–frequency dual–polarization applications.

III. Design of dual–polarization dual–frequency reflectarray prototype

A dual–polarization dual–frequency focused beam reflectarray antenna has been designed for transmit–receive (Tx–Rx) operation in Ku Band. The reflectarray element employed consists of two orthogonal sets of three parallel dipoles, and it has been characterized in the previous section. The reflectarray has been designed to radiate a main beam in the direction $\theta_b = 16.9^\circ$ and $\varphi_b = 0^\circ$ at the Tx center frequency of 11.95 GHz (see Fig. 3). The reflectarray is circular and consists of 861 elements arranged in a 33×33 grid with cell size $12\,\mathrm{mm} \times 12\,\mathrm{mm}$. The phase center of the feed-horn is assumed to be located at the point of coordinates $x = -193$ mm, $y = 0$ mm and $z = 635$ mm with respect to the center of the reflectarray (see Fig. 3). The radiation pattern of the feed-horn is modeled as a function $\cos^q(\theta)$ [1], where q is varied with frequency to provide an illumination level at the reflectarray edges of roughly 10 dB below the maximum. The geometrical dimensions of the elements have been adjusted by means of the implemented spectral domain MoM software. In particular, the phase of the reflection coefficient for each element has been matched to the required phase that produces a collimated beam as described in [1].

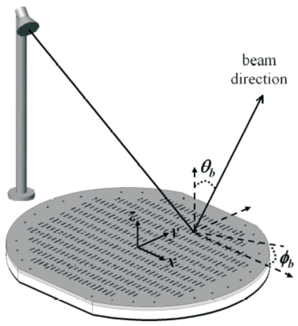
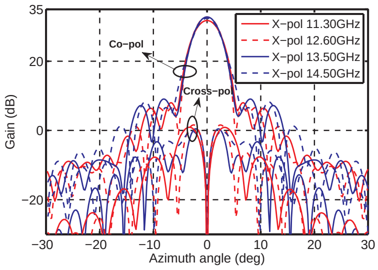
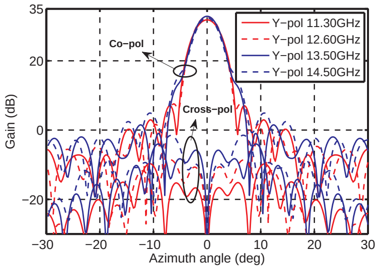


Fig. 3. Coordinate system for a generic reflectarray antenna made of elements with two orthogonal sets of three parallel dipoles. The angular coordinates θ_b and φ_b indicate the main beam direction for a collimated beam reflectarray.

We have observed that the use of orthogonal dipoles of the same size in Fig. 1 leads to an important increase in the cross–polar component of the reflected electric field. This effect is due to coupling between the orthogonal dipoles. Fortunately, the coupling can be substantially reduced by using in each cell orthogonal dipoles with very different dimensions, which also reduces the cross–polarized electric field. Therefore, in the design of the reflectarray of Fig. 3, we have imposed a phase shift of 180 degrees between the required phase of the reflection coefficient for the X–polarization and that required for the Y–polarization. According to Fig. 2, the phase range is larger than 360 degrees in all the cases, making possible to select orthogonal dipoles with very different dimensions, and consequently, lower cross–polarization in the resulting antenna.

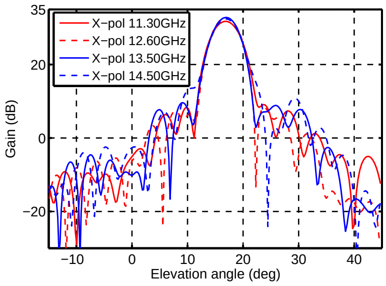


(a)

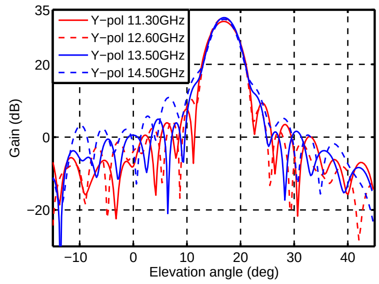


(b)

Fig. 4. Radiation patterns in azimuth plane for a dual–polarization dual–frequency reflectarray antenna based on the element with two orthogonal sets of three parallel dipoles shown in Fig. 1. Co–polar and cross–polar components are shown. (a) Results for X–polarization. (b) Results for Y–polarization.



(a)



(b)

Fig. 5. Radiation patterns in elevation plane for a dual–polarization dual–frequency reflectarray antenna based on the element with two orthogonal sets of three parallel dipoles shown in Fig. 1. Only the co–polar component is shown. (a) Results for X–polarization. (b) Results for Y–polarization.

Figs. 4(a) and 4(b) show the co–polar and cross–polar radiation patterns of the designed reflectarray in the azimuth plane at the limits of both the Tx frequency interval (11.3–12.6 GHz: 10.9% bandwidth) and the Rx frequency interval (13.5–14.5 GHz: 7.1% bandwidth). The gain variations are less than 1 dB in these two frequency intervals, and the radiation patterns are reasonably well preserved. The sidelobes are roughly kept 25 dB below the maximum, and the maximum cross–polar radiation is about 30 dB below the main beam co–polar radiation, which indicates that the reflectarray element based on orthogonal dipoles of different dimensions is a low cross–polarization element.

Fig. 5 shows the co–polar radiation patterns in the elevation plane at the limits of both the Tx frequency interval and the Rx frequency interval. Again, we can see that the gain variations are less than 1 dB in the two frequency intervals, and that the sidelobes are roughly kept 25 dB below the maximum in the whole bands. The cross–polar radiation is not shown in Figs. 5(a) and 5(b) because it is 60dB below the main beam co–polar radiation (remind the elevation plane is a symmetry plane). We would like to point out that numerical optimization tools that enforce the required phase in the two frequency bands [9] hasn't been required. This indicates that the reflectarray element of Fig. 1 is a broadband element.

IV. Improvement of Cross–Polar Discrimination by Means of Element Rotation

We can note that the cross–polar component of the radiation patterns for Y–polarization in Fig. 4(b) is much lower than the cross–polar component for X–polarization in Fig. 4(a). However, if we compute the cross–polar components for both polarizations at the design frequency 11.95 GHz by assuming that the dipoles don't produce cross–polar fields (see the dashed lines in Fig. 6), we can note that the cross–polar component of the radiation pattern for X–polarization is much lower than the cross–polar component for Y–polarization. The ideal cross–polar components (those obtained when the cross–polar fields of the dipoles are neglected) and the real cross–polar components (those obtained when the cross–polar fields of the dipoles are considered) switch their roles as shown in Fig. 6. In the case of the Y–polarization, what happens is that the contribution of the dipoles to the cross–polarization cancels out the contribution of the feed-horn to the cross–polarization when both contributions are accounted for simultaneously (blue solid line in Fig. 6). However, in the case of the X–polarization, the contribution of the feed-horn to the cross–polarization is negligible with respect to that of the dipoles, and the resulting cross–polarization (red solid line in Fig. 6) is mainly due to the dipoles. Therefore, if we want to reduce the cross–polarization in the case of the X–polarization, we should act on the dipoles. One possible way to do this is to introduce a slight rotation of the dipoles inside the unit cell as depicted in Fig. 7. In order to reduce the cross–polarization of the reflectarray antenna analyzed in Figs. 4 and 5 in the case of the X–polarization, we have introduced slight rotations in the reflectarray elements within ± 5 degrees. An optimization routine has been used which minimizes the contribution of the cross–polar electric field of each element to the cross–polarization (in the X–polarization) as a function of the rotation angle (provided this angle is within ± 5 degrees). This optimization routine calls a MoM–SD code that rigorously analyzes the rotated element in a periodic environment. The final results obtained for the radiation patterns of the reflectarray antenna with rotated elements are shown in Fig. 8. We can see that there is a reduction of the cross–polar component for X–polarization with respect to the cross–polar level shown in Fig. 5 at the cost of a small increase of the cross–polar component for Y–polarization. These results show that a tradeoff betweeen the level of the two cross-polar componenets can be achieved. Note that in Fig. 8 the cross–polar radiation for the two polarizations is always 35 dB below the main beam co–polar radiation, which indicates that an improvement of cross–polar discrimination can be achieved by slight rotations of the elements of the reflectarray.

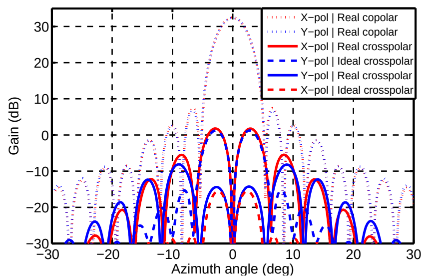


Fig. 6. Radiation patterns in the azimuth plane for the designed reflectarray at 11.95 GHz in two cases: the reflected cross–polar electric fields of the dipoles are neglected (ideal) and the reflected cross–polar fields of the dipoles are considered (real).

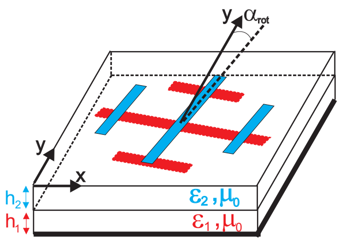


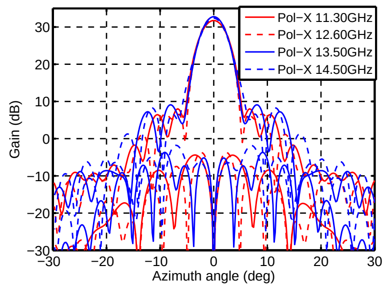
Fig. 7. Rotation of the reflectarray element based on the two orthogonal sets of three parallel dipoles.

V. Conclusions

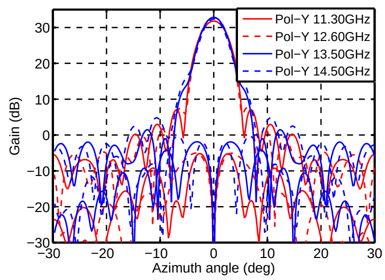
A dual–polarization dual–frequency focused beam reflectarray antenna for transmit–receive (Tx–Rx) operation in Ku Band has been designed. The reflectarray element consists of two orthogonal sets of three parallel dipoles placed at the two sides of a dielectric layer. The reflectarray shows a bandwidth of 10.9% in the Tx-band and 7.1% the in Rx-band for a gain variation lower than 1 dB. In order to improve cross–polar discrimination in both polarizations, the elements of the designed reflectarray have been rotated within ±5 degrees, As a result, a level of cross-polarization 35 dB below the main beam copolar radiation has been achieved for both linear polarizations. These results show that the proposed reflectarray can be used to realize terrestrial and satellite transmit/receive antennas for space communications in Ku-band.

Acknowledgment

This work has been supported by by the Spanish Ministry of Science and Innovation (projects CICYT TEC2010-17567 and CONSOLIDER-INGENIO CSD2008-68), by European Space Agency under contract 4000106334, and by Junta de Andalucía (project P09–TIC–4595).



(a)



(b)

Fig. 8. Radiation patterns in azimuth plane for the dual–polarization dual–frequency reflectarray antenna analyzed in Figs. 4 and 5 when the elements are slightly rotated within ±5 degrees. Co–polar and cross–polar components are shown. (a) Results for X–polarization. (b) Results for Y–polarization.

References

[1] J. Huang and J. A. Encinar, *Reflectarray antennas*. Piscataway, NJ/New York: IEEE Press/Wiley, 2008.

[2] D.M. Pozar, S.D. Targonski, and H.D. Syrigos, “Design of millimeter wave microstrip reflectarray,” *IEEE Trans. Antennas Propagat.*, vol. 45, No. 2, pp. 287-296, February 1997.

[3] M. Bozzi, S. Germani, L. Perregrini, “Performance comparison of different element shapes used in printed reflectarrays,” *Antennas and Wireless Propagation Letters*, Volume 2, Issue 1, 2003 pp. 219 - 222.

[4] J. A. Encinar, “Design of two-layer printed reflectarrays using patches of variable size,” *IEEE Trans. Antennas Propagat.*, vol. 49, pp. 1403–1410, Oct. 2001.

[5] J.A. Encinar, A. Pedreira, “Flat reflector antenna in printed technology with improved bandwidth and separate polarizations”, Granted patent P2004 01382.

[6] R. Florencio, R. R. Boix, E. Carrasco, J. A. Encinar and V. Losada, “Efficient numerical tool for the analysis and design of reflectarrays based on cells with three parallel dipoles”, *Microw. Opt. Technol. Lett.*, accepted for publication, 2013.

[7] R. Florencio, R.R Boix, J.A. Encinar, “Efficient analysis of multi-resonant periodic structures for the improved analysis and design of reflectarray antennas”, *Proc. 6th Edition EUCAP*, Prague, Czech Republic, March 2012.

[8] J.A. Encinar, et al. "Dual-polarisation reflectarray antenna with improved cros-polarization properties" European patent, EP2337152 (A1), 10/12/2009.

[9] J. A. Encinar, L. Datashvili, J. A. Zornoza, M. Arrebola, M. Sierra-Castañer, J. L. Besada, H. Baier, and H. Legay, “Dual-polarization dual-coverage reflectarray for space spplications,” *IEEE Trans. on Antennas and Propag.*, Vol. 54, No. 10, Pp. 2828-2837, Oct. 2006.

[10] R. Mittra, C. H. Chan, and T. Cwik, “Techniques for analyzing frequency selective surfaces-A review”,*Proc. IEEE.* vol. 76, no. 12, pp. 1593-1615, Dec. 1988.